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Advanced control structures for induction motors with ideal current loop response using field oriented control

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ABSTRACT

Field oriented control (FOC) is widely used for high performance induction motor (IM) electrical drive systems. Typically, FOC uses linear controls and space vector modulation (SVM) to control the fundamental components of the stator voltages. This work shows that based on a fast and precise inner current loop response one may flexibly employ different advanced control methods, to achieve high performance outer loops (speed and flux control). In this paper, novel approaches based on dead-beat scheme for the current loop combining with exact linearization, backstepping controls, and fatness-based methods for the outer loop are proposed. By comparing with classical PI control, the proposed method shows the outstanding features of system response such as fast, accurate and decoupling properties. The performance evaluation is given by experimental results.

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1. INTRODUCTION

Nowadays, asynchronous electrical drives based on field-oriented control (FOC) have been widely used in industrial applications [1]. Based on this method, we can find induction motors have similar characteristics to separate excitation DC motors in term of generating magnetic field and torque [1-4]. In the FOC structure, when the stator voltage control satisfies the requirement of "fast – accuracy – decoupling" properties in current response, the induction motor can be considered as fed by a current source inverter with controllable current, which leads to order reduction of the model of induction motor drive from 4^{th} to 2^{th} order [1].

The article presents different methods to design the inner loop (stator current loop) and outer loop (flux and speed loops). Firstly, deadbeat control with finite response is employed for the current loops [5-7]. Secondly, exact linearization is utilized to transform the non-linear dynamics of current model into linear input-output relationship, thus it is able to apply common linear controls to the current model [8, 9]. In addition, to demonstrate current performances, a classical PI current controller is also designed [10-13] for benchmarking purpose. The closed-loop current response based on deadbeat, exact linearization control is evaluated to identify the most suitable control for the current loops. The success in designing a control for the current loops can lead to the assumption of ideal current loop response that results in the system order reduction. Subsequently, the design control of the electrical drive system outer loops based on reducing order model can be performed in various methods. The classical PI controller can only be effective around the operating point. When operating in a wide range the system performance can be degraded [10-13].

Nowadays, the non-linear control with abrupt developments in hardware are increasingly considered in practical applications.

Due to flat property of the IM with rotor speed and flux are selected as the outputs, the flatness-based principle is used for speed and flux control. By simply reducing the order of the governing equations, the designed speed and flux reference trajectories can be choosing based on the amplitude constraint of current [14-18]. Noting that the IM model is of strict feedback form, backstepping control method [19-21] which makes sure that the error between set values and real values satisfy Lyapunov's stability is also deployed for speed and flux loops. Evaluation of dynamic response between different speed and flux control structures based on ripple torque performance [22-25].

The advanced structures for FOC of the induction motor with ideal current loop are verified according experimental results. These results are obtained from the assessment of current loop response, speed, flux and the performance of electrical drive between FOC structures such as harmonic distortion, ripple torque, and max ripple torque. The remainder of this paper is organized as follows. The mathematical model of the drive system with ideally control performance of the stator system will be presented in section 2. Subsequently, the design method of stator current control and outer loop by nonlinear method is discussed in section 3 and 4. The efficiency of the proposed method is verified by simulation as well as implementation are shown in section 5 and section 6, respectively. The final section will summarize the research and gives some directions for future works

2. MATHEMATICAL MODEL OF THREE PHASE INDUCTION MOTOR

In synchronous coordinate, the three-phase induction motor can be described by the following dynamical [1].

$$\begin{cases} \dfrac{di_{sd}}{dt} = -\left(\dfrac{1}{\sigma T_s} + \dfrac{1-\sigma}{\sigma T_r}\right) i_{sd} + \omega_s i_{sq} + \dfrac{1-\sigma}{\sigma T_r} + \dfrac{1}{\sigma L_s} u_{sd} \\ \dfrac{di_{sq}}{dt} = -\omega_s i_{sd} - \left(\dfrac{1}{\sigma T_s} + \dfrac{1-\sigma}{\sigma T_r}\right) i_{sq} - \dfrac{1-\sigma}{\sigma} \omega i_m + \dfrac{1}{\sigma L_s} u_{sq} \\ \dfrac{d\psi_{rd}}{dt} = -\dfrac{1}{T_r}\psi_{rd} + \dfrac{L_m}{T_r} i_{sd} \\ \dfrac{d\omega}{dt} = k_\omega \psi_{rd} i_{sq} - \dfrac{z_p}{J} m_L \end{cases} \tag{1}$$

With $\omega_s = \omega + \omega_r = \omega + \dfrac{L_m}{T_r}\dfrac{i_{sq}}{\psi_{rd}}; k_\omega = \dfrac{3}{2}\dfrac{z_p^2 L_m^2}{L_r J}$

In which, $i_{sd}; i_{sq}$ are dq components of the stator current; i_m is electromagnetic currents; ω, ω_s are mechanical and synchronous speed, respectively; ψ'_{rd}, ψ'_{rq} are dq components of the rotor flux; σ is total leakage factor; T_r is rotor time constant; u_{sd}, u_{sq} are dq components of the stator voltage; L_s is stator inductance, m_L : torque load; m_W : torque motor. It can be seen that the original state (1) is bilinear and is of 4th order. When considering the current controller response is perfect, the induction motor model can be reduced as:

$$\begin{cases} \dfrac{d\psi_{rd}}{dt} = -\dfrac{1}{T_r}\psi_{rd} + \dfrac{L_m}{T_r} i_{sd} \\ \dfrac{d\omega}{dt} = k_\omega \psi_{rd} i_{sq} - \dfrac{z_p}{J} m_L \end{cases} \tag{2}$$

The state (2) is of 2nd order, stator current i_{sd} is used to control the motor flux and i_{sq} is dedicated to speed control

3. STATOR CURRENT CONTROL DESIGN

3.1. Deadbeat control

The ideal dynamic behavior can be achieved by deadbeat response which means that the actual value will track the reference value after one sampling period, or, if the one-step delay of the control output is taken into account, after two sampling periods. This leads to the fact that closed loop transfer function must also have the form of a polynomial of Nth degree, with the sum of polynomial coefficients equal to [1,5,6]. According to [1], the problem of designing controller transfer function is now replaced by finding a polynomial of matrix controller. Where the polynomial $L(z^{-1})$ has to fulfill:

$$L(1)=\frac{1}{B(1)} \rightarrow \sum_{i=0}^{s} l_i = 1\Big/\sum_{j=0}^{m} b_j \tag{3}$$

With: l_i coefficients of the polynomial $L(z^{-1})$; B numerator of process transfer functions $G_{S.}$
b_j coefficients of the polynomial $B(z^{-1})$, numerator of $G_S(z^{-1})$
$L(1)$, $B(1)$ sum of polynomial coefficients of $L(z^{-1})$, $B(z^{-1})$
According to to [1,5,6] the stator current control with deadbeat behavior as shown as:

$$\mathbf{R}_I(z)=\begin{bmatrix} \dfrac{(z-\Phi_{11})L_1(z^{-1})}{1-z^{-1}L_1(z^{-1})} & \dfrac{-\Phi_{12}L_2(z^{-1})}{1-z^{-1}L_2(z^{-1})} \\ \dfrac{\Phi_{12}L_1(z^{-1})}{1-z^{-1}L_1(z^{-1})} & \dfrac{(z-\Phi_{11})L_2(z^{-1})}{1-z^{-1}L_2(z^{-1})} \end{bmatrix} \tag{4}$$

Where: $\Phi_{11};\Phi_{12}$ transition matrix; T : sampling time ; T_s is stator time constant.

$$\Phi_{11}=\begin{vmatrix} 1-\dfrac{T}{\sigma}\left(\dfrac{1}{T_s}+\dfrac{1-\sigma}{T_r}\right) & 0 \\ 0 & 1-\dfrac{T}{\sigma}\left(\dfrac{1}{T_s}+\dfrac{1-\sigma}{T_r}\right) \end{vmatrix};\ \Phi_{11}=\begin{vmatrix} \dfrac{1-\sigma}{\sigma}\dfrac{T}{T_r} & \dfrac{1-\sigma}{\sigma}\omega T \\ -\dfrac{1-\sigma}{\sigma}\omega T & \dfrac{1-\sigma}{\sigma}\dfrac{T}{T_r} \end{vmatrix}$$

It can be deduced that the control laws do not cause future errors (stationary control errors), the polynomials $L_1(z^{-1})$ and $L_2(z^{-1})$ must not contain the coefficient l_0. Additionally, to eliminate the stationary control errors the transfer function $\mathbf{G}_w$ of the closed loop must be equal $\mathbf{I}$ under stationary conditions ($z=1$). Therefore, it can be seen that:

$$L_1(1)=L_2(1)=1 \tag{5}$$

With the conditions (5) we cannot clearly define the coefficients of polynomials $L_1(z^{-1})$ and $L_2(z^{-1})$ when only the total coefficient polynomials is by 1. Therefore, it is necessary to have sub-conditions to be able to determine those coefficients, which is the condition of voltage constraint. Selection of $L_1(z^{-1})$ and $L_2(z^{-1})$ as second-degree polynomial as follows

$$L_1(z^{-1})=L_2(z^{-1})=l_1 z^{-1}+l_2 z^{-2} \tag{6}$$

The controller (6) can be written in form of discrete equations:

$$\begin{cases} y_d(k)=l_1 y_d(k-2)+l_2 y_d(k-3)+l_1 e_d(k)+(l_2-l_1\Phi_{11})e_d(k-1) \\ \qquad -l_2\Phi_{11}e_d(k-2)-l_1\Phi_{12}e_q(k-1)-l_2\Phi_{12}e_q(k-2) \\ y_q(k)=l_1 y_q(k-2)+l_2 y_q(k-3)+l_1 e_q(k)+(l_2-l_1\Phi_{11})e_q(k-1) \\ \qquad -l_2\Phi_{11}e_q(k-2)+l_1\Phi_{12}e_d(k-1)+l_2\Phi_{12}e_d(k-2) \end{cases} \tag{7}$$

The controller (7) can be written in the following form:

$$\begin{cases} u_{sd}(k+1) = h_{11}^{-1}\left[y_d(k) - \Phi_{13}\psi'_{rd}(k+1)\right] \\ u_{sq}(k+1) = h_{11}^{-1}\left[y_q(k) + \Phi_{14}\psi'_{rd}(k+1)\right] \end{cases} \tag{8}$$

Where l_1 and l_2 can be chosen as follows:

$$l_1 = \min\left\{\frac{h_{11}u_{d0}}{e_{d0}}; \frac{h_{11}u_{q0} - \Phi_{14}i_{sdN}}{e_{q0}}\right\}; l_2 = 1 - l_1 \tag{9}$$

With: $h_{11} = \dfrac{T}{L_{sd}}$ input matrix; $\Phi_{14} = \dfrac{1-\sigma}{\sigma}\omega T$.

The control structure of the inner loop using the dead-beat is shown in Figure 1:

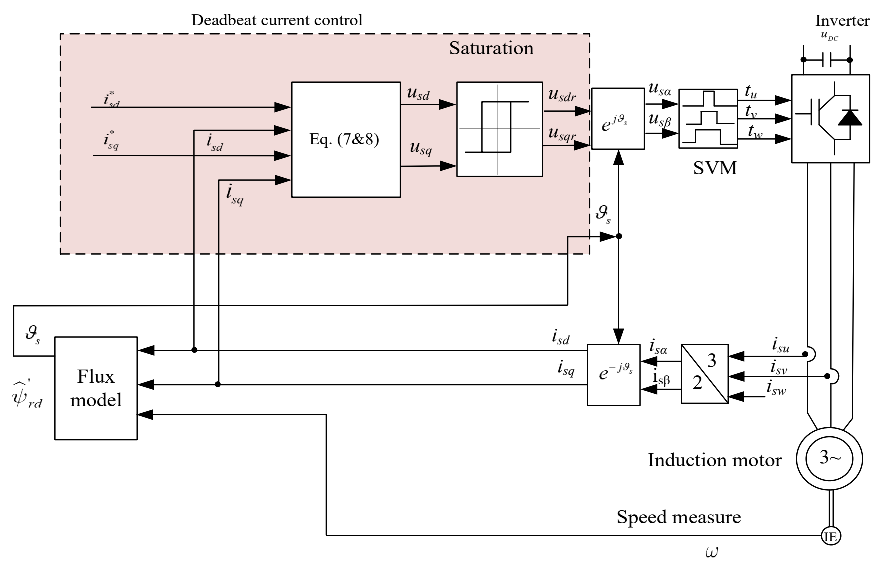

Figure 1. Control structure of the current vector controller with deadbeat

3.2. Exact linearization control method

Using the state feedback or the coordinate transformation the exact linearized IM model can be represented. The transformed state model will now become the starting point for the controller design. Besides the exact linearization achieved in discrete state space, the input-output decoupling relations are totally guaranteed. Based on this result, it seems to be possible to replace the two dimensional current controller by a coordinate transformation and two separate current controllers for both *d* and *q* axes [1,8,9]. According to [1,8,9] the state-feedback control law can be written as:

$$\mathbf{u} = -\mathbf{L}^{-1}(\mathbf{x})\mathbf{g}(\mathbf{x}) + \mathbf{L}^{-1}(\mathbf{x})\mathbf{w} = \mathbf{a}(\mathbf{x}) - \mathbf{L}^{-1}(\mathbf{x})\mathbf{w} \tag{10}$$

The (10) can be written in detailed form:

$$\mathbf{u}=\begin{bmatrix}\dfrac{dx_1}{a}-\dfrac{c\psi'_{rd}}{a}\\ \dfrac{dx_2}{a}+\dfrac{cT_r\omega\psi'_{rd}}{a}\\ 0\end{bmatrix}+\begin{bmatrix}\dfrac{1}{a} & 0 & \dfrac{-x_2}{a}\\ 0 & \dfrac{1}{a} & \dfrac{x_1}{a}\\ 0 & 0 & 0\end{bmatrix}\begin{bmatrix}\mathrm{w}_1\\ \mathrm{w}_2\\ \mathrm{w}_3\end{bmatrix} \tag{11}$$

With: parameters: $a=\dfrac{1}{\sigma L_s};c=\dfrac{(1-\sigma)}{\sigma T_r};d=\dfrac{1}{\sigma T_s}+\dfrac{(1-\sigma)}{\sigma T_r};w_1=u_{sd};w_2=u_{sq};w_3=\vartheta_s$

The state-feedback control law or the coordinate transformation law can be written in detailed form:

$$\begin{cases}\mathrm{w}_1=u_{sd}=\dfrac{1}{a}[i_{sd}+c\dfrac{\psi_{rd}}{L_m}+\mathrm{w}_1-i_{sq}\mathrm{w}_3-i_{sq}\vartheta_s]\\ \mathrm{w}_2=u_{sq}=\dfrac{1}{a}[i_{sq}-cT_r\omega\dfrac{\psi_{rd}}{L_m}+\mathrm{w}_2+i_{sd}\mathrm{w}_3+i_{sd}\vartheta_s]\\ \mathrm{w}_3=\omega_s\end{cases} \tag{12}$$

The control structure of the inner loop using the method of exact linearization is shown in Figure 2

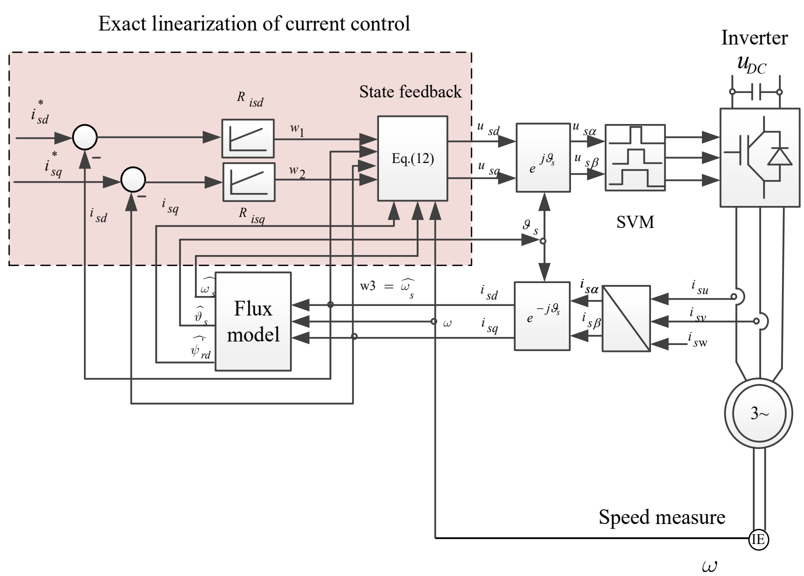

Figure 2. Control structure of the current vector controller with method of exact linearization

4. NONLINEAR CONTROL DESIGN FOR OUTER LOOPS

Since the deadbeat controller which forces the stator current to its desired value in finite steps has already successfully been developed for the stator current loop in previous sections, the remaining objective of this research is to design controller for torque and flux loops.

4.1. Flatness-based control design

The asynchronous electrical drive structure using flatness-based control for the outer loop (speed and flux loop) with ideal performance of stator current is shown in Figure 4. Based on [1,17], the flat output is the flux and speed $\left[\psi_r^d, \omega^d\right]$,thus we obtain the flatness-based controller for the flux and speed as follows:

- *Flux and speed reference trajectory designing*

In order to guarantee differentiable property, the relationship between $\psi_{rd}^* \& \psi_{rd}^d$ and $\omega^* \& \omega$ can be given as follows:

$$G_1 = \frac{\psi_{rd}^*(s)}{\psi_{rd}^d(s)} = \frac{1}{1 + T_1 s + T_1^2 s^2} \tag{13}$$

$$G_2 = \frac{\omega^*(s)}{\omega(s)} = \frac{1}{1 + T_2 s + T_2^2 s^2} \tag{14}$$

With: ψ_{rd}^*, ω^* : reference flux and speed; ψ_{rd}^d, ω flux and speed reference trajectory

- *Feedforward control designing*

In the flat system, the input value is $u_T = \left[i_{sd}^*,\ i_{sq}^*\right]$, the input value $i_{sd}^*; i_{sq}^*$ are computed based on the flat output and the constructed trajectories:

$$i_{sd}^* = \psi_{rd}^* + T_r \frac{d\psi_{rd}^*}{dt} \tag{15}$$

$$i_{sq}^* = \frac{h_1 \frac{d\omega^*}{dt} + \tilde{m}_L}{h_2 \psi_{rd}^*} \tag{16}$$

With $\tilde{m}_L$: estimate torque load

- *Feedback control designing*

In fact, the model state of IM is not absolutely accurate, and the disturbances that have negative effects on the quality of the flatness-based controller. For this reason, in order to ensure that the output match the desired value, additional closed loop control is required using PI controller. Therefore, the PI controller for the speed and flux loop can be expressed as (17) and (18):

$$R_\psi(z) = V_\psi \frac{(1 - d_\psi . z^{-1})}{(1 - z^{-1})} \tag{17}$$

Where: $V_\psi \approx \dfrac{1}{3\left(1 - e^{-T_\psi / T_r}\right)}$ $d_\psi \approx e^{-T_\psi / T_r}$

And

$$G_r\left(z^{-1}\right) = \frac{V_R\left(1 + d_1 . z^{-1}\right)}{\left(1 - z^{-1}\right)} \tag{18}$$

Where:

$d_1 = a_1$: parameters; $V_R = \dfrac{\left(1 + z_2\right)^2}{V_s\left[\left(1 + 3m\right) + z_2\left(-1 + m\right)\right]}$: amplification coefficient

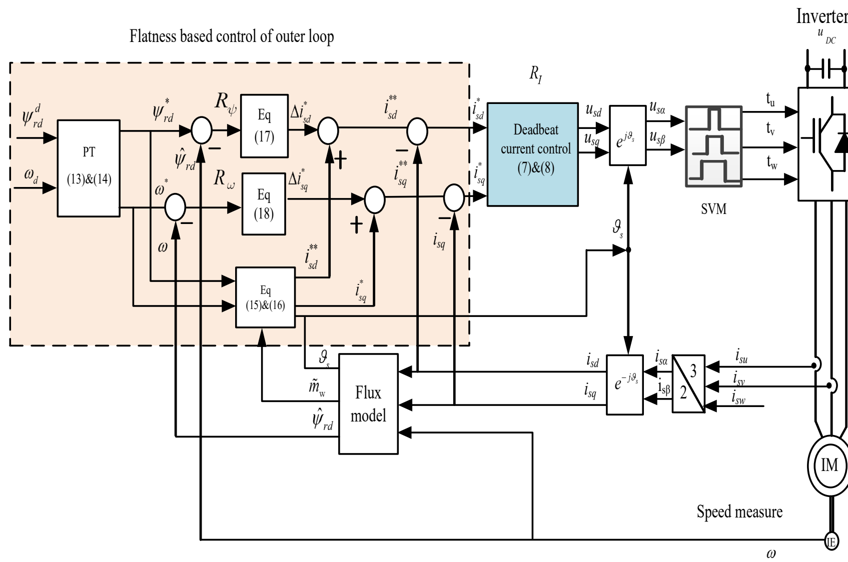

Figure 3. FOC control structure with the deadbeat controller for the IM' stator current loop and the flatness-based controller for the outer loop

4.2. Backstepping control design

Backstepping approach is based on feedback controller designing that satisfies Lyapunov's stability by constructing CLFs (control Lyapunov functions) from subsystems [19-21]. The asynchronous electrical drive structure using backstepping control method for the outer loop with ideally control performance of stator current is shown in Figure 4.

- *Flux control loop designing*

According to [19] using the backstepping control method, the flux controller is expressed as follows:

$$i_{sd} = -c_1 T_r z_1 + \psi_{rd} + T_r \frac{d\psi^*_{rd}}{dt} \tag{19}$$

With : c_1 : constant; $z_1 = \psi_{rd} - \psi^*_{rd}$: errors of rotor flux; $\psi^*_{rd};\psi_{rd}$: reference and actual flux

With the ideal control performance of stator current $i^*_{sd} \approx i_{sd}$, one can obtain:

$$i^*_{sd} = -c_1 T_r z_1 + \psi_{rd} + T_r \frac{d\psi^*_{rd}}{dt} \tag{20}$$

Where i^*_{sd} is the actual control signal of the flux controller.

- *Speed control loop designing*

Similarly, the speed controller is as follows:

$$i^*_{sq} = \frac{1}{k}(-c_2 z_2 + \frac{z_p}{J} m_w + \frac{d\omega^*}{dt}) \tag{24}$$

With : c_2 : constant; $z_2 = \omega^* - \omega$: errors of speed; $\omega^*;\omega$: reference and actual speed

With the ideal control performance of stator current $i^*_{sq} \approx i_{sq}$, that leads to:

$$i_{sq} = \frac{1}{k}(-c_2 z_2 + \frac{z_p}{J} m_L + \frac{d\omega^*}{dt}) \tag{25}$$

- *Design of set point trajectory for flux and speed*

In addition to the goal of making the set point reference enough differentiable like output, the constraints of output could be taken into account during trajectory design as well. The flux and speed reference trajectory designing by (13) and (14).

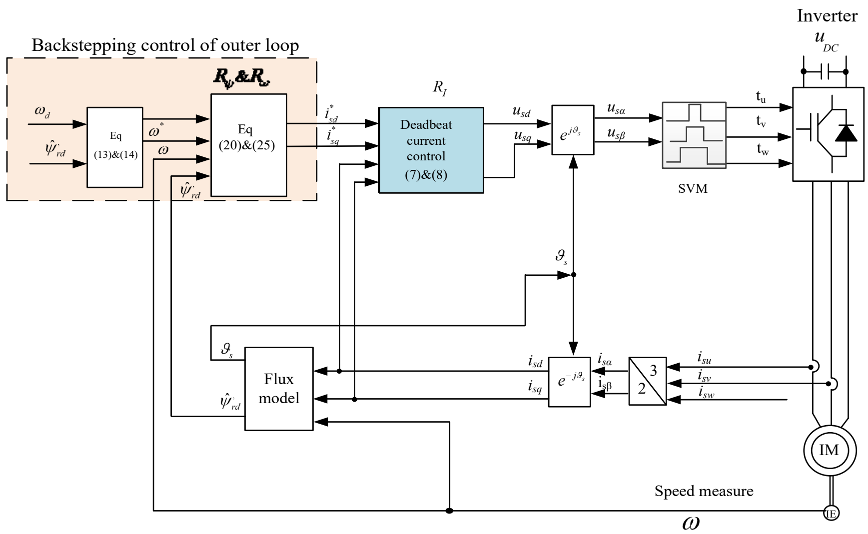

Figure 4. FOC control structure with the deadbeat controller for the IM' stator current loop and the backstepping controller for the outer loop

5. EXPERIMENTAL RESULTS

In order to evaluate the effectiveness of stator current control and speed control methods, the induction motor is operated with IM's parameters as shown in Table 1:

Table 1. Experimental with IM's parameters

IM parameters	Symbol	Value
Power	P_{dm}	2.2 kW
Rated speed	N_{dm}	2880 rpm
Rated voltage	U_{dm}	400V
Pole pair	Z_p	1
Power factor	$cos\varphi$	0.9
Inertia torque	J	0.002Kg.m2

This section gives out the evaluation of experimental results that includes stator current response, speed, flux response and performance of electrical drive.

The evaluation of the magnetization of the squirrel cage induction motor will be done by taking the finite sampling step as simulation which are shown in Figure 5 and Table 2.

Table 2. Performance comparison of the stator current loop

FOC control structures	Exact linearization		Deadbeat	
Currents	i_{sd}	i_{sq}	i_{sd}	i_{sq}
Settling time (s)	0.0025	0.0025	0.001	0.001
Overshoot (%)	15	25	10	20

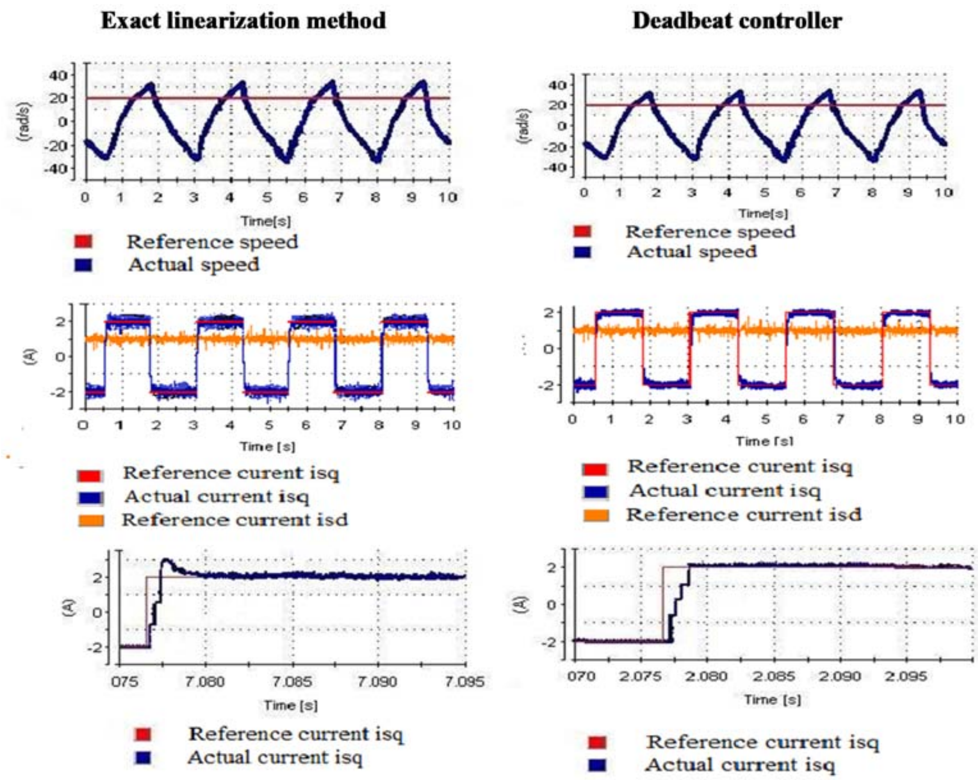

Figure 5. Dynamical responses of the stator currents loop

This are results in the Figure 5 and Table 2, we can the the current controller with the exact linearization method has the large overshoot and ripple current i_{sq}. Besides, the currents i_{sd}, i_{sq} of deadbeat controller satisfies “fast – accuracy – decoupling” properties compared to the exact linearization that leads fast speed response and low ripple.

When the torque is required (reference torque is match with actual torque), followed by the paper presented for dynamical response evaluation of FOC control structures at speed of 0.1 rad/s and 100 rad/s are expressed through Figure 6, Figure 7 and Table 3:

Table 3. System response comparison

FOC control structures	Deadbeat-Flatness	Deadbeat- Backstepping
	at the speed of 0.1 rad/s	
Flux settling time (s)	0.4	0.35
Speed settling time (s)	0.05	0.05
	at the speed of 100 rad/s	
Flux settling time (s)	0.3	0.25
Speed settling time (s)	0.25	0.15

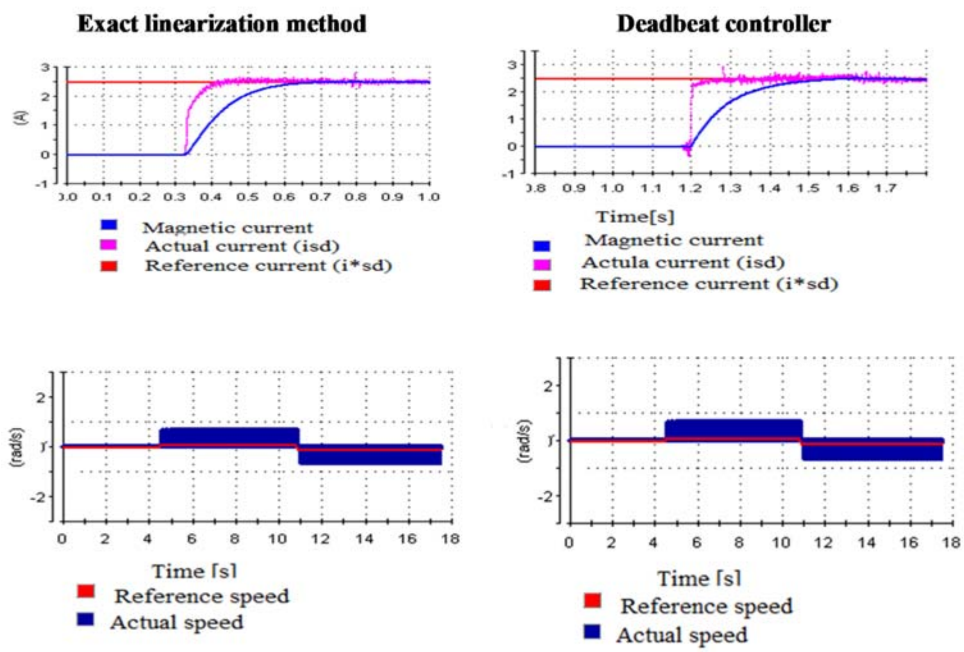

Figure 6. i_{sd} current and speed responses at the speed of 0.1 rad/s

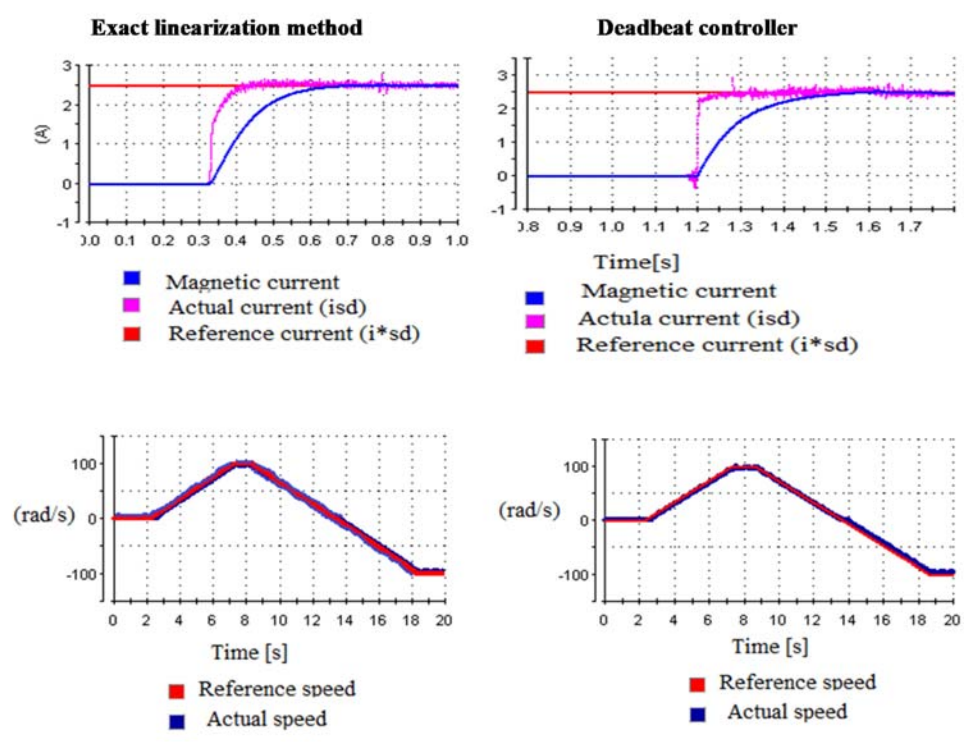

Figure 7. i_{sd} current and speed responses at the speed of 100 rad/s

Based on the above results, we can the flux settle within 0.25s, the long over speed time with wide speed range, overshoot about 10% to 25%. It is worth noting that the outer loop control using backtepping method has short settling time compared to the other responses. In depth evaluation of proposed control are expressed through Figure 8, Figure 9 and Table 3.

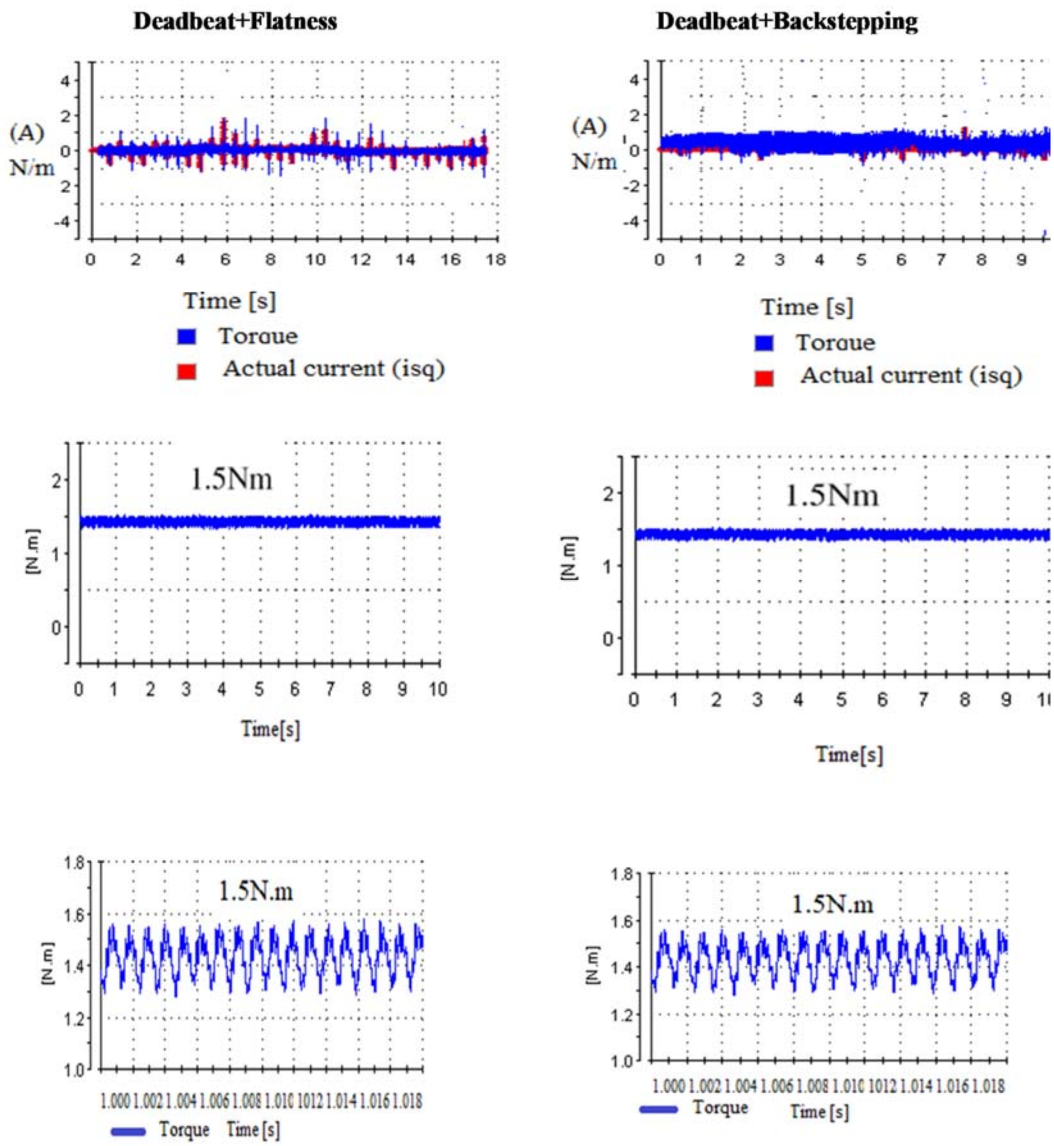

Figure 8. Torque responses at the speed of 0.1 rad/s

Table 3. Experimental test results of the FOC control structures

FOC control structures	Deadbeat-Flatness	Deadbeat-Backstepping
	at the speed of 0.1 rad/s	
Maximum ripple torque-no load (ΔT_m%)	50	30
Ripple torque –loaded (RT_F%)	7.0	6.9
	at the speed of 100 rad/s	
Maximum ripple torque - no load (ΔT_m%)	8	5
Ripple torque –loaded (RT_F%)	9.3	9.25

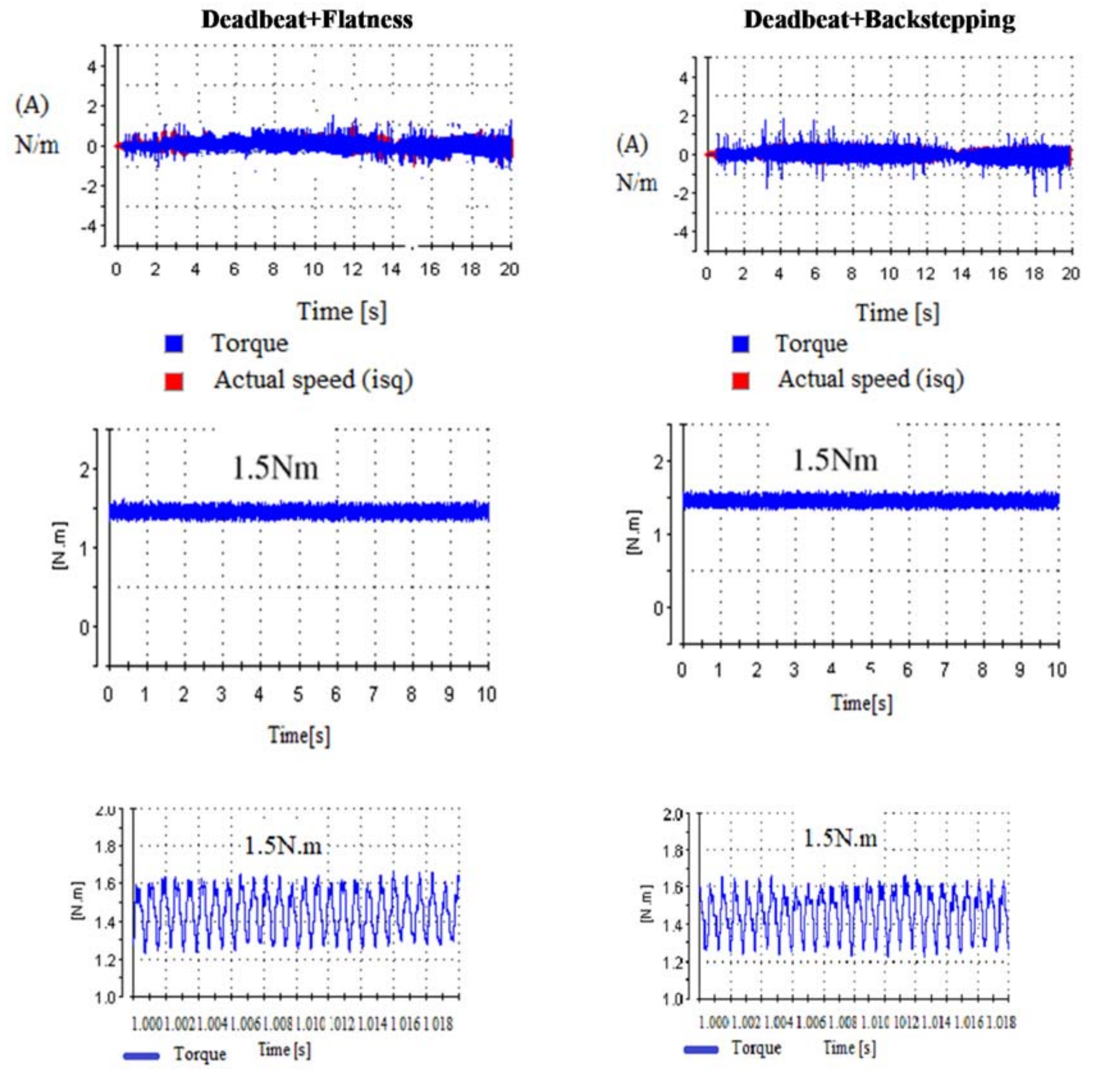

Figure 9: Torque responses at the speed of 100 rad/s

Based on the above results, we can the maximum ripple torque-no load ΔT_m% from 30% to 50%, the ripple torque –loaded (RT_F%) within 7% at the speed of 0.1 rad/s. The maximum ripple torque-loaded ΔT_m% from 5% to 8%, the ripple torque –loaded (RT_F%) within 9% at the speed of 100 rad/s. It can be observed from numerical and experimental results that backstepping control method provides better performances than those of the other two controls.

6. CONCLUSION

In the paper, we introduced the stator current loop controller based on deadbeat and backstepping control-methods for the outer loop deliver with some advances in term of kinetic response and qualified electrical drive. The accuracy of the proposed methods is demonstrated by experiment. The results show that the deadbeat controller has already successfully been developed for the stator current loop, satisfying the requirement of "fast – accuracy – decoupling". The research results also suggest some ways to design the outer loop control for complicated drive systems where the motor is coupled with varying load via flexible coupling.

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